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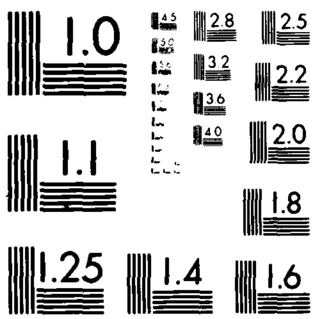

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FOREIGN TECHNOLOGY DIVISION

AIR FORCE SYSTEMS COMMAND

ANTENNA

(Selected Articles)

FOREIGN TECHNOLOGY DIVISION

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FTD-ID(RS)T-0993-81

EDITED TRANSLATION

FTD-ID(RS)T-0993-81 2 November 1981

MICROFICHE NR: FTD-81-C-001000

ANTENNA (Selected Articles)

English pages: 25

Source: Antenny, Nr. 8, 1970, pp. 28-46

Country of origin: USSR
Translated by: Bernard L. Tauber
Requester: FTD/TQFE
Approved for public release; distribution unlimited.

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А а	*А а*	A, a	Р р	*Р р*	R, r
Б б	*Б б*	B, b	С с	*С с*	S, s
В в	*В в*	V, v	Т т	*Т т*	T, t
Г г	*Г г*	G, g	У у	*У у*	U, u
Д д	*Д д*	D, d	Ф ф	*Ф ф*	F, f
Е е	*Е е*	Ye, ye; E, e*	Х х	*Х х*	Kh, kh
Ж ж	*Ж ж*	Zh, zh	Ц ц	*Ц ц*	Ts, ts
З з	*З з*	Z, z	Ч ч	*Ч ч*	Ch, ch
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М м	*М м*	M, m	Ь ь	*Ь ь*	'
Н н	*Н н*	N, n	Э э	*Э э*	E, e
О о	*О о*	O, o	Ю ю	*Ю ю*	Yu, yu
П п	*П п*	P, p	Я я	*Я я*	Ya, ya

*ye initially, after vowels, and after ъ, ь; e elsewhere. When written as ё in Russian, transliterate as yё or ё.

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

Russian	English	Russian	English	Russian	English
sin	sin	sh	sinh	arc sh	$\sinh^{-1}$
cos	cos	ch	cosh	arc ch	$\cosh^{-1}$
tg	tan	th	tanh	arc th	$\tanh^{-1}$
ctg	cot	cth	coth	arc cth	$\coth^{-1}$
sec	sec	sch	sech	arc sch	sech^{-1}
cosec	csc	csch	csch	arc csch	csch^{-1}

Russian	English
rot	curl
lg	log

COMPENSATION FOR DIRECT AND REVERSE INTERCONNECTION IN AN INFINITE EMITTER ARRAY

L. V. Ryzhkova

An infinite emitter array is presented in the form of a multiterminal network whose input connectors coincide with the inputs of actual emitters and whose output connectors coincide with the outputs of individual emitters. This permits expressing all elements of the emitter array scatter matrix by the coupling coefficient between the input connectors. The article studies the possibility of compensation for the leakage of energy from one of the channels to the inputs of other channels and the reradiation, by the apertures, of the energy which has landed on them from other apertures.

Introduction

The presence of an interconnection in phased arrays leads to an entire series of unpleasant consequences which are manifested especially sharply when scanning in a broad sector. This, primarily, is the mismatch of emitters with the feeding lines and the distortion of the radiation pattern which is caused by the disruption of amplitude-phase distribution which is given by the control system. Thus, the task of weakening the influence of the interconnection on the operation of the antenna is extremely important.

A method for calculating the parameters of a circuit to compensate for reverse interconnection in arbitrary linear arrays was proposed in [1]. Here we present a more general method which permits compensating for the mismatch (reverse interconnection) as well as the reradiation of the energy of each emitter by other apertures (direct interconnection).

Matrix of the Scatter of an Infinitely Long Emitter Array

Let us present an infinitely long emitter array in the form of a multiterminal network whose input connectors coincide with the inputs of actual emitters and output connectors - with the outputs of "individual emitters" [2]. In this case, the system of waves reflected from the output connectors of the antenna is a system of amplitude-phase distributions in the aperture of the antenna, each of which is formed with the excitation of one elementary emitter while all the other emitters are closed on matched loads. The radiation patterns of individual emitters, which would coincide with the radiation patterns of elementary emitters if there were no interconnection between the emitters of the system, correspond to these amplitude-phase distributions in the aperture.

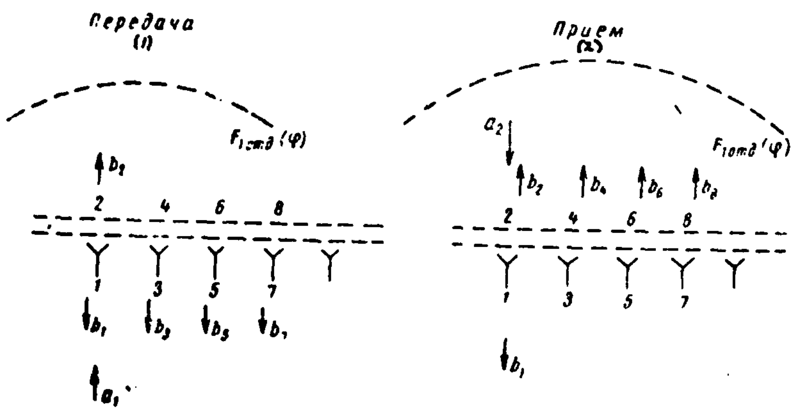

Fig. 1.
KEY: (1) Transmission; (2) Reception.

With the presence of an interconnection, the excitation of one input causes the appearance of reflected waves on all other inputs and distortion the radiation pattern of the excited emitter due to the waves which are reradiated by all the other apertures (Fig. 1).

Let us write the scatter matrix of the multiterminal network which has been introduced in the following form:

$$\left[\begin{array}{cccccccc|cccccccc}
 & \vdots & \vdots & \vdots & \vdots & & & & & \vdots & \vdots & \vdots & \vdots & \vdots & & \\
\ldots & S_{11} & S_{13} & S_{15} & S_{17} & \ldots & & & \ldots & S_{12} & 0 & 0 & 0 & 0 & \ldots & \\
\ldots & S_{13} & S_{11} & S_{13} & S_{15} & \ldots & & & \ldots & 0 & S_{12} & 0 & 0 & 0 & \ldots & \\
\ldots & S_{15} & S_{13} & S_{11} & S_{13} & \ldots & & & \ldots & 0 & 0 & S_{12} & 0 & 0 & \ldots & \\
\ldots & S_{17} & S_{15} & S_{13} & S_{11} & \ldots & & & \ldots & 0 & 0 & 0 & S_{12} & 0 & \ldots & \\
 & \vdots & \vdots & \vdots & \vdots & & & & & \vdots & \vdots & \vdots & \vdots & & & \\
\hline
 & \vdots & \vdots & \vdots & \vdots & & & & & \vdots & \vdots & \vdots & \vdots & & & \\
\ldots & S_{12} & 0 & 0 & 0 & \ldots & & & \ldots & S_{22} & S_{24} & S_{26} & S_{28} & \ldots & & \\
\ldots & 0 & S_{12} & 0 & 0 & \ldots & & & \ldots & S_{24} & S_{22} & S_{24} & S_{26} & \ldots & & \\
\ldots & 0 & 0 & S_{12} & 0 & \ldots & & & \ldots & S_{26} & S_{24} & S_{22} & S_{24} & \ldots & & \\
\ldots & 0 & 0 & 0 & S_{12} & \ldots & & & \ldots & S_{28} & S_{26} & S_{24} & S_{22} & \ldots & & \\
 & \vdots & \vdots & \vdots & \vdots & & & & & \vdots & \vdots & \vdots & \vdots & & &
\end{array}\right]$$

Here the inputs are numbered by the numbers 1, 3, 5 ..., and the outputs - 2, 4, 6 ... . The elements of the scatter matrix have the following meaning: S_{11} - the reflection coefficient on the input of some emitter with the excitation of it alone; S_{13} - reflection coefficient at the input of some emitter with the excitation of the adjacent one alone (i.e., "leakage" - to the input of the adjacent emitter); S_{15} - "leakage" across one and so forth; S_{12} - the transmission factor from some input to its own output which determines the amplitude-phase distribution in the aperture which corresponds to the radiation pattern of an individual emitter, i.e., to the radiation pattern of the entire antenna with the excitation of only one input; S_{22} - the scatter coefficient of an individual emitter; S_{24} - the connection of two adjacent outputs ("reradiation") and so forth.

With this definition of the elements of a scatter matrix, all types of reflections and scattering in the system prove to be considered. The following relationships are valid here:

$$\left.\begin{array}{l}\sum\limits_{k=1,3\ldots}^{\infty} |s_{1,k}|^2 + |s_{1,2}|^2 = 1 \\ s_{1,k} s_{1,2}^{*} + s_{1,2} s_{2,k+1}^{*} = 0 \\ \sum\limits_{k=1,3\ldots}^{\infty} |s_{2,k+1}|^2 + |s_{1,2}|^2 = 1\end{array}\right\}, \tag{1}$$

which express the unitary nature of the scatter matrix.

Using the system of equations (1) we can express all elements of the scatter matrix through the elements of its first quadrant which determine the connection of the input connectors:

$$\left.\begin{array}{l}s_{1,2} = \sqrt{1 - \sum\limits_{k=1,3,\ldots}^{\infty} |s_{1,k}|^2}\, e^{i\varphi_{12}} \\ s_{2,k+1} = -s_{1,k}^{*} e^{i2\varphi_{12}}\end{array}\right\}. \tag{2}$$

The elements of the first quadrant $S_{1,k}$ can be determined from the matrix of impedances Z of the antenna from the formula

$$\begin{bmatrix} & \vdots & \vdots & \vdots & \\ \ldots & s_{11} & s_{13} & s_{15} & \ldots \\ \ldots & s_{13} & s_{11} & s_{13} & \ldots \\ \ldots & s_{15} & s_{13} & s_{11} & \ldots \\ & \vdots & \vdots & \vdots & \end{bmatrix} = (Z-1)(Z+1)^{-1}.$$

or measured. Thus, all elements of the scatter matrix can be determined.

Now, let us select from the infinite array one emitter with incident and reflected waves a_1, b_1, a_2, b_2 at the input and output connectors. We can write formally that these waves are connected to each other using some matrix

$$S_A = \begin{bmatrix} s_{I\,I} & s_{I\,II} \\ s_{II\,I} & s_{II\,II} \end{bmatrix}.$$

If the entire array is excited uniformly and cophasally, the amplitude of the reflected wave at the input of the emitter being examined, with consideration of leakage from all the other emitters, is determined by the coefficient

$$s_{II} = \sum_{k=1,3\ldots}^{\infty} s_{1,k}.$$

The transmission coefficient of such an emitter

$$s_{I\,II} = s_{II\,I} = s_{1,2}.$$

The scatter of energy by the emitter aperture is determined by the coefficient

$$s_{II\,II} = \sum_{k=1,3\ldots}^{\infty} s_{2,k+1}.$$

As follows from (2)

$$s_{II\,II} = -s^{*}_{II}\, e^{i2\varphi_{12}}.$$

Then we can write the scatter matrix of an emitter which is operating in the system in the following form

$$S_A = \begin{bmatrix} \sum\limits_{k=1,3\ldots}^{\infty} s_{1,k} & s_{1,2} \\ s_{1,2} & -\sum\limits_{k=1,3\ldots}^{\infty} s^{*}_{1,k}\, e^{i2\varphi_{1,2}} \end{bmatrix}.$$

Thus, when the antenna operates on transmission the four-terminal network forms an incident wave on the input connectors of one emitter a_1 to wave b_2 in the aperture, the amplitude and phase of which determine the pattern of the given individual emitter in a distant zone. Here, added to the reflected wave itself at the input is the wave which leaks through from all other emitters to the input of the element being considered. With the operation of the antenna on receive, the wave in aperture a_2 which corresponds to the radiation pattern of a given separate emitter during operation on transmission is transformed by a four-terminal network into wave b_1 which is received at the input of the given emitter. Here, added to the scattered wave itself at the output are waves which landed on the aperture being examined from other apertures and are reradiated by them.

In this examination, no consideration is given to the fraction of the energy which arrives from the element being examined at the inputs of all other elements and the energy from its aperture which is reradiated by other apertures. Therefore, it can be considered that S_A is the scatter matrix of some hypothetical device which possesses losses. On the otherhand, since in the general case

$$\sum_{k=1,3,\ldots}^{\infty} |s_{1,k}|^2 \neq \left| \sum_{k=1,3,\ldots}^{\infty} s_{1,k} \right|^2 ,$$

it is clear that matrix S_A does not satisfy the relationships of unitarity. The introduction of such a matrix is formal procedure which subsequently permits the easy calculation of the interesting parameters of a real system.

In conclusion of this section, let us move on to a transmission matrix of a single emitter which operates in the system

$$T_A = \begin{bmatrix} \dfrac{s_{1,2}^2 + \Sigma\Sigma^* e^{i2\varphi_{12}}}{s_{1,2}} & -\dfrac{\Sigma^* e^{i2\varphi_{12}}}{s_{1,2}} \\ -\dfrac{\Sigma}{s_{1,2}} & \dfrac{1}{s_{1,2}} \end{bmatrix},$$

where

$$\Sigma = \sum_{k=1,3\ldots}^{\infty} s_{1,k}.$$

Now the problem should consist of compensating for the reflections to the input of the emitter due to the "reverse" interconnection ("leakage") or the scatter of energy into space due to direct interconnection ("reradiation").

Scatter Matrix of an Infinitely Long Array of Waveguides Connected by Openings

Let an array consist of an infinitely large number of identical waveguides with electrical length $\mathscr{E}$ connected by openings at a distance $\mathscr{E}/2$ from the input (Fig. 2). We write the scatter matrix of such a system in the following form:

$$\left[\begin{array}{cccccc|cccccc}
 & \vdots & \vdots & \vdots & \vdots & & & \vdots & \vdots & \vdots & \vdots & \\
\cdots & \Gamma & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \cdots & \cdots & \sqrt{1-\Gamma^2-\beta^2} & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \cdots \\
\cdots & \frac{ia}{2} & \Gamma & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \cdots & \cdots & \frac{ia}{2} & \sqrt{1-\Gamma^2-\beta^2} & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \cdots \\
\cdots & \left(\frac{ia}{2}\right)^2 & \frac{ia}{2} & \Gamma & \frac{ia}{2} & \cdots & \cdots & \left(\frac{ia}{2}\right)^2 & \frac{ia}{2} & \sqrt{1-\Gamma^2-\beta^2} & \frac{ia}{2} & \cdots \\
\cdots & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^2 & \frac{ia}{2} & \Gamma & \cdots & \cdots & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^2 & \frac{ia}{2} & \sqrt{1-\Gamma^2-\beta^2} & \cdots \\
 & \vdots & \vdots & \vdots & \vdots & & & \vdots & \vdots & \vdots & \vdots & \\
\hline
 & \vdots & \vdots & \vdots & \vdots & & & \vdots & \vdots & \vdots & \vdots & \\
\cdots & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^4 & \left(\frac{ia}{2}\right)^5 & \cdots & \cdots & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^4 & \left(\frac{ia}{2}\right)^5 & \cdots \\
\cdots & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^4 & \cdots & \cdots & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \left(\frac{ia}{2}\right)^4 & \cdots \\
\cdots & \sqrt{1-\Gamma^2-\beta^2} & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \cdots & \cdots & \Gamma & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \left(\frac{ia}{2}\right)^3 & \cdots \\
\cdots & \frac{ia}{2} & \sqrt{1-\Gamma^2-\beta^2} & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \cdots & \cdots & \frac{ia}{2} & \Gamma & \frac{ia}{2} & \left(\frac{ia}{2}\right)^2 & \cdots \\
 & \vdots & \vdots & \vdots & \vdots & & & \vdots & \vdots & \vdots & \vdots &
\end{array}\right]$$

Here $\Gamma e^{i\gamma}$ - the coefficient of reflection from the opening at the input; $\frac{i\alpha}{2} e^{i\varphi}$ - the transmission coefficient to the input of the adjacent waveguide; $\left(\frac{i\alpha}{2}\right)^2 e^{i\varphi}$ - the coefficient of transmission to the waveguide through one, etc.; $\sqrt{1-\Gamma^2-\beta^2}\, e^{i\delta}$ - the coefficient of transmission from the input to the output of one waveguide.

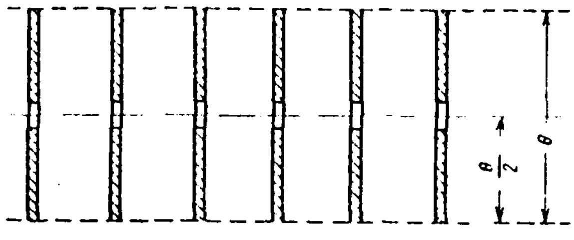

Fig. 2.

From the relationships which express the unitarity of the scatter matrix of an infinite array of waveguides, it is necessary to determine the relationships between the elements Γ, α, and β. The condition which expressed the energy balance in the system:

$$\Gamma^2 + (1 - \Gamma^2 - \beta^2) + 4\sum_{k=1}^{\infty}\left(\frac{\alpha^2}{4}\right)^k = 1.$$

Hence we obtain

$$\beta^2 = 4\sum_{k=1}^{\infty}\left(\frac{\alpha^2}{4}\right)^k.$$

Summing this series, we obtain

$$\beta^2 = \frac{4\alpha^2}{4-\alpha^2}; \qquad (3)$$

Let us now write the condition for orthogonality of any two lines of the matrix between which there are n lines where n = 0, 1, 2, ..., in the following form:

$$\sum_{m=1}^{5} Q_m = 0; \qquad (4)$$

where the following are designated:

$$Q_1 = \Gamma\left[\left(\frac{i\alpha}{2}\right)^{n+1}\right]^* + \Gamma\left[\left(\frac{i\alpha}{2}\right)^{n+1}\right];$$

$$Q_2 = \sqrt{1-\Gamma^2-\beta^2}\left[\left(\frac{i\alpha}{2}\right)^{n+1}\right]^* + \sqrt{1-\Gamma^2-\beta^2}\left[\left(\frac{i\alpha}{2}\right)^{n+1}\right];$$

$$Q_3 = 2\sum_{k=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k\left[\left(\frac{i\alpha}{2}\right)^{k+n+1}\right]^*;$$

$$Q_4 = 2\sum_{k=1}^{\infty}\left(\frac{i\alpha}{2}\right)^{k+n+1}\left[\left(\frac{i\alpha}{2}\right)^k\right]^*;$$

$$Q_5 = 2\sum_{k=1}^{n}\left(\frac{i\alpha}{2}\right)^k\left[\left(\frac{i\alpha}{2}\right)^{k+n+1}\right]^*.$$

It is easy to show that

$$Q_1 + Q_2 = \begin{cases} 0 & \text{with n even,} \\ \left(\Gamma + \sqrt{1-\Gamma^2-\beta^2}\right)\left[\pm 2\left(\frac{\alpha}{2}\right)^{n+1}\right] & \text{with n odd;} \end{cases}$$

$$Q_3 + Q_4 = \begin{cases} 0 & \text{with n even,} \\ \frac{2\alpha^2}{4-\alpha^2}\left[\pm 2\left(\frac{\alpha}{2}\right)^{n+1}\right] & \text{with n odd;} \end{cases}$$

$$Q_5 = \begin{cases} 0 & \text{with n even,} \\ \mp 2\left(\frac{\alpha}{2}\right)^{n+1} & \text{with n odd;} \end{cases}$$

(With the presence of a double sign in the formula with even n's the upper sign pertains to n's which are multiples of 4 and the lower - to n's which are not multiples of 4; with uneven n's the upper sign pertains to (n + 1) multiples of 4 and the lower - to (n + 1) which are not multiples of 4.)

Now it is obvious that with even n's condition (4) is always satisfied.

With odd n's

$$\sum_{m=1}^{5} Q_m = \pm 2\left(\frac{\alpha}{2}\right)^{n+1}\left[\Gamma + \sqrt{1-\Gamma^2-\beta^2} + \frac{2\alpha^2}{4-\alpha^2} - 1\right].$$

Consequently, in order to satisfy the condition of orthogonality of the lines it is necessary that

$$\Gamma + \sqrt{1-\Gamma^2-\beta^2} + \frac{2\alpha^2}{4-\alpha^2} - 1 = 0. \qquad (5)$$

Taking (3) into account, from (5) we obtain

$$\Gamma + \sqrt{1-\Gamma^2-\beta^2} + \frac{\beta^2}{2} - 1 = 0. \qquad (6)$$

Since $\beta^2>0$, in solving (6) relative to β^2 we find:

$$\beta^2 = 2\sqrt{\Gamma}(\sqrt{2-\Gamma} - \sqrt{\Gamma}).$$

Thus, the relationship between elements of the scatter matrix of an infinite array of waveguides without losses has been determined.

Next we perform the same operation as with an infinite array of emitters.

Let us select one waveguide and, for it, we write down the non-unitary scatter matrix

$$S_M = \begin{bmatrix} s'_{I\,I} & s'_{I\,II} \\ s'_{II\,I} & s'_{II\,II} \end{bmatrix}.$$

The coefficient of reflection at the input

$$s'_{I\,I} = s'_{II\,II} = \left[\Gamma + 2\sum_{k=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k\right]e^{i\zeta}.$$

Summing this series we obtain

$$s'_{I\,I} = \left(\Gamma + \frac{2i\alpha}{2-i\alpha}\right)e^{i\zeta}.$$

The transmission coefficient after summing the series

$$s'_{II\,I} = s'_{I\,II} = \left(\sqrt{1-\Gamma^2-\beta^2} + \frac{2i\alpha}{2-i\alpha}\right)e^{i\zeta}.$$

Thus, the scatter matrix of a waveguide in the system

$$S_M = e^{i\zeta}\begin{bmatrix} \Gamma + \frac{2i\alpha}{2-i\alpha} & \sqrt{1-\Gamma^2-\beta^2} + \frac{2i\alpha}{2-i\alpha} \\ \sqrt{1-\Gamma^2-\beta^2} + \frac{2i\alpha}{2-i\alpha} & \Gamma + \frac{2i\alpha}{2-i\alpha} \end{bmatrix}.$$

Changing to the transmission matrix, we obtain

$$T_M = \begin{bmatrix} \frac{(B_2+iC_2)^2-(B_1+iC_1)^2}{B_2+iC_2}\,e^{i\zeta} & \frac{B_1+iC_1}{B_2+iC_2} \\ -\frac{B_1+iC_1}{B_2+iC_2} & \frac{1}{B_2+iC_2}\,e^{-i\zeta} \end{bmatrix},$$

where

$$B_1 = \Gamma - \frac{2\alpha^2}{4-\alpha^2};\; B_2 = \sqrt{1-\Gamma^2-\beta^2} - \frac{2\alpha^2}{4-\alpha^2};\; C_1 = C_2 = \frac{4\alpha}{4+\alpha^2}.$$

Conditions for Compensating for Interconnection

We connect in cascade four-terminal networks which correspond to an emitter and waveguide which operate in infinite arrays. The transmission matrix of such a connection is determined by the expression

$$T' = T_A T_M$$

Hence we obtain elements of a compound transmission matrix:

$$t'_{I\,I} = \frac{s_{12}^2 - \Sigma\Sigma^* e^{i2\varphi_{11}}}{s_{12}} \frac{(B_2 + iC_2)^2 - (B_1 + iC_1)^2}{B_2 + iC_2} e^{i\theta} + \frac{\Sigma^* e^{i2\varphi_{11}}}{s_{12}} \frac{B_1 + iC_1}{B_2 + iC_2};$$

$$t'_{I\,II} = \frac{s_{12}^2 - \Sigma\Sigma^* e^{i2\varphi_{11}}}{s_{12}} \frac{B_1 + iC_1}{B_2 + iC_2} - \frac{\Sigma^* e^{i2\varphi_{11}}}{s_{12}} \frac{1}{B_2 + iC_2} e^{-i\theta};$$

$$t'_{II\,I} = -\frac{\Sigma}{s_{12}} \frac{(B_2 + iC_2)^2 - (B_1 + iC_1)^2}{B_2 + iC_2} e^{i\theta} - \frac{1}{s_{12}} \frac{B_1 + iC_1}{B_2 + iC_2};$$

$$t'_{II\,II} = -\frac{\Sigma}{s_{12}} \frac{B_1 + iC_1}{B_2 + iC_2} + \frac{1}{s_{12}} \frac{1}{B_2 + iC_2} e^{-i\theta}.$$

In order to write the conditions for compensation for interconnection without changing to a compound scatter matrix of a cascade connection, we turn to the general formulas of the transition T ->S [3]:

$$s'_{11} = -\frac{t'_{II\,I}}{t'_{II\,II}};\quad s'_{12} = \frac{1}{t'_{II\,II}};\quad s'_{21} = \frac{t'_{II}\,t'_{II\,II} - t'_{I\,II}\,t'_{II\,I}}{t'_{II\,II}};\quad s'_{22} = \frac{t'_{I\,II}}{t'_{II\,II}}.$$

For the connection not to have reflections on the input ($s'_{11} = 0$), it is necessary that $t'_{III} = 0$; so that there is no scatter ($s'_{22} = 0$) it is necessary that $t'_{I\,II} = 0$.

From the first condition it follows that

$$\Sigma e^{i\theta}[(B_2 + iC_2)^2 - (B_1 + iC_1)^2] + (B_1 + iC_1) = 0.$$

Hence we can obtain the following formula which connects the elements of the scatter matrix of the emitter array with the parameters of a waveguide system which compensates for the "reverse" interconnection

$$|\Sigma|^2 = \frac{\left(\Gamma - \frac{2\alpha^2}{4+\alpha^2}\right)^2 + \frac{16\alpha^2}{(4+\alpha^2)^2}}{\left(1 - 2\Gamma^2 - \frac{4\alpha^2}{4-\alpha^2} - \frac{4\alpha^2\sqrt{1-\Gamma^2-\beta^2}}{4+\alpha^2} - \frac{4\alpha\Gamma}{4+\alpha^2}\right)^2 \rightarrow}$$

$$\rightarrow \frac{}{+ \frac{64\alpha^2(\sqrt{1-\Gamma^2-\beta^2} - \Gamma)^2}{(4+\alpha^2)^2}}$$

Disregarding the small values of Γ^2, α^4, $\Gamma\alpha^2$, we obtain the approximate compensation condition

$$|\Sigma|^2 \simeq \frac{2\alpha^2}{2+\alpha^2},$$

whence we obtain a convenient calculation formula to determine the transmission coefficient of an aperture which connects the waveguides which form the compensation circuit

$$\alpha_1 \simeq \sqrt{\frac{2|\Sigma|^2}{2-|\Sigma|^2}}. \qquad (7)$$

To compensate the "direct" interconnection we use the condition

$$T_{I\,II} = 0,$$

from which, after conversion, we obtain the precise

$$\Gamma^2 + \frac{4\alpha^2(1-\Gamma)}{4+\alpha^2} = \frac{|\Sigma|^2}{|s_{12}|^2 + |\Sigma|^2}$$

and the approximate

$$\alpha_2 = \sqrt{\frac{4|\Sigma|^2}{4(|s_{12}|^2 + |\Sigma|^2) - |\Sigma|^2}} \qquad (8)$$

conditions for compensation of reradiation.

Now the values of electrical length of waveguides can easily be obtained for circuits for compensating for leakage and reradiation

$$\operatorname{tg}\theta_1 = \frac{-k_1 2\alpha(2\Gamma-1) + k_2(3\alpha^2 - 2\Gamma)}{k_1(24\alpha^2 + 16\Gamma) + k_2 16\alpha(2\Gamma-1)}, \qquad (9)$$

$$\operatorname{tg}\theta_2 = \frac{k_1 4\alpha - k_2(4\Gamma - 2\alpha^2)}{k_1(4\Gamma - 2\alpha^2) - k_2 4\alpha}, \qquad (10)$$

where

$$\Sigma = k_1 + i k_2.$$

Conclusion

As follows from formulas (7)-(10), in an infinite array of emitters where there are only two controllable parameters with the selected method of compensation at our disposal - the coefficient of transmission of the compensating opening and its electrical distance from the aperture, only the separate compensation for energy leakage from the aperture of each emitter to the inputs of the remainder and reradiation of energy which landed on it from other apertures by each aperture is possible in the general case.

In this work, the entire examination was conducted for the case of the uniform and cophasal excitation of the emitters. However, it can be shown that the compensation circuits whose parameters are determined by the formulas obtained will also operate with the scanning of a beam, i.e., with the non-cophasal excitation of the emitters. This statement is based on the fact that with the examined method of compensation each channel which consists of an emitter and waveguide with openings seemingly itself compensates for its influence on the other channels; therefore, the phase of a wave which lands, for example, from the emitter with number i in the channel with number (i + 1) (Fig. 3) changes with scanning due to the connection of the aperture in the same way as the phase of a wave which lands from the i-th channel to the (i + 1)-th through a compensating opening.

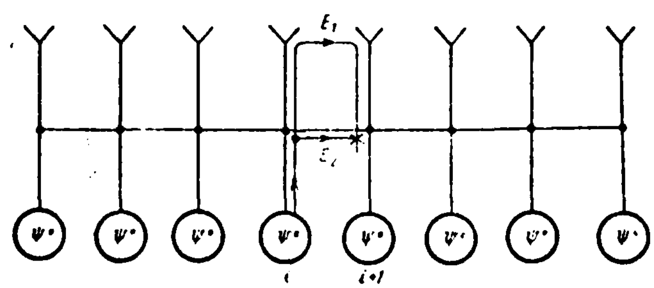

Fig. 3.

The formulas which have been obtained can be used to accomplish the decoupling of emitters in large linear arrays.

In conclusion, I consider it my duty to express my profound graditude to O. G. Vendik for constant attention to the work.

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SIMULTANEOUS COMPENSATION OF DIRECT AND REVERSE INTERCONNECTION IN AN INFINITE EMITTER ARRAY

L. V. Ryzhkova

A method is proposed for the calculation of a scheme to compensate for the interconnection of emitters of a plane array in the form of a system of openings which connect the feeding waveguide of all emitters of the array with each other. It is shown that the simultaneous compensation for the leakage of energy from one of the channels to the inputs of other channels (reverse interconnection) and the reradiation by apertures of the energy which landed on them from other apertures (direct interconnection) is possible in a two-dimensional array.

Formulation of the Problem

In a plane array of emitters, it is possible to have available four controlled parameters instead of two, as occurred in a linear antenna, virtually almost without complicating the system [1]. Therefore, the possibility appears here for the simultaneous compensation for leakage of energy from the aperture of each emitter to the inputs of the remainder and the reradiation, by each aperture, of the energy which landed on it from other apertures.

We can introduce in the plane array a system of openings in the narrow and broad walls of the feeding waveguides for the simultaneous compensation of the connection between lines and columns of the array.

The task of this work is to clarify the conditions under which the simultaneous compensation of direct and reverse interconnection is possible and to determine the parameters of the compensation circuit.

We can proceed in the following manner to solve the problem of compensation for interconnection in a plane array: examine separately the connection of the emitters within the line and then the connection between the lines. From the conditions of compensation for connection within the line a determination is made of the parameters of the openings in the narrow walls for a unified emitter-line as was done for a linear antenna [1]. Then, based on the conditions for the compensation of the connection between lines or between unified emitters-lines, a determination is made of the parameters of the openings in the broad walls for a unified emitter of the array, i.e., for an emitter-column which consists of emitter-lines, again just as for a linear antenna. Separate compensation in planes E and H was used by Hannan [2] who first proposed the application of compensating lines of connection between emitters to match the elements of phased arrays with feeding lines.

In our case, the task is posed somewhat more broadly - it is necessary to achieve not only the matching of elements with the feeding lines, i.e., compensation for reflected waves on the input connectors of the system of emitters (reverse interconnection), but also to exclude the scattering of energy into space, consequently, compensating for reradiated waves (direct interconnection). Therefore, a somewhat different approach which consists of the following appears to be more promising here. A unified element-array connected with all the other elements of the array is examined. Then, a unified element of a compensating waveguide array which is also connected with all the other waveguides is connected to this emitter. Next we find the compound scatter matrix of a cascade connection of the array element with the element of the waveguide circuit and equations of compensation whose solution will provide the parameters of the connection openings are compiled.

In the conclusion of this section, we should note that, first, each isolated element of the emitter array is considered ideally matched with its feeding line and the mismatch and distortion of the radiation pattern which arise with the operation of such an emitter in the system are considered; second, an infinite array is introduced for simplification since in it all elements are identical and are under the same environmental conditions.

Scatter Matrix of an Element of Emitter Array

Let us examine the structure of a scatter matrix of an infinite array

$$S' = \left[\begin{array}{c|c} S'_{II} & S'_{I\,II} \\ \hline S'_{II\,I} & S'_{II\,II} \end{array}\right].$$

Here S'_{II} is the scatter matrix for the inputs whose elements are: the reflection coefficient of the array element itself; the connection of the array element with all elements of its line; the correction of the array element with all elements of its column; the connection of the array with the elements of all other columns, excluding those elements which belong to a line already selected (i.e., all connections are considered and no connection is repeated twice).

$S'_{II\,I}$, $S'_{I\,II}$ - are identical units which characterize the transmission from the input of one element to all outputs including its own; if the radiation pattern of a "separate emitter" is considered as an output, as in the case of a linear antenna [3], then these blocks are diagonal matrices.

$S_{II\,II}$ is the matrix of scatter of the system among the outputs whose elements are the scatter coefficient of the element and the connection between the radiation pattern of individual emitters.

Just as formerly,

$$S'_{I\,I} = (Z - 1)(Z + 1)^{-1}.$$

where Z - the matrix of array impedences, I - a single matrix. We now prepare a scatter matrix of an emitter in the system

$$R = \begin{bmatrix} R_{11} & R_{12} \\ R_{21} & R_{22} \end{bmatrix}.$$

We determine the coefficient of reflection on input R_{11} as the sum of the reflection coefficient itself and of all coefficients of transmission which characterize the leakage of energy to the input of a given emitter from all the others, i.e., as the sum of all elements of a line of block S'_{II}:

$$R_{11} = s_{11} + \sum_{k=3,5\ldots}^{\infty} s_{1k}. \qquad (1)$$

We determine the coefficient of reflection on output R_{22} as the sum of the scatter coefficient itself and of all transmission coefficients which characterize the reradiation of the energy of all other apertures by a given emitter:

$$R_{22} = s_{22} + \sum_{k=4,6\ldots}^{\infty} s_{2k}.$$

The transmission coefficient of a unitized emitter is determined by the diagnonal element of block S'_{III}:

$$R_{12} = R_{21} = |s_{12}|\, e^{i\varphi_{12}}.$$

Obviously, just as in the case of an infinite array, we can show that

$$\underset{k=1,3\ldots}{s'_{2,k+1}} = -S^{*}_{1k}\, e^{i2\varphi_{12}}.$$

Then

$$R_{22} = -R^{*}_{11}\, e^{i2\varphi_{12}}.$$

From the condition of unitarity of the scatter matrix of an infinite array

$$|s_{12}| = \sqrt{1 - \sum_{k=1,3\ldots}^{\infty} |s_{1k}|^2}.$$

Then

$$|R_{12}| = \sqrt{1 - \sum_{k=1,3\ldots}^{\infty} |s_{1k}|^2}. \quad (2)$$

Thus, if the elements of the input block of the scatter matrix of the array are measured or calculated, then all elements of the scatter matrix of the emitter in the system can be determined. In moving on to the transmission matrix of this arbitrary emitter, we have

$$T' = \begin{bmatrix} \frac{R_{21}R_{12} - R_{22}R_{11}}{R_{12}} & \frac{R_{22}}{R_{12}} \\ -\frac{R_{11}}{R_{12}} & \frac{1}{R_{12}} \end{bmatrix}.$$

Scatter Matrix of a Unified Element of a Compensation Circuit

Let us first turn to an infinite array which is formed by waveguides with electrical length Θ which are connected with each other by openings in narrow (transmission coefficient of opening α, reflection coefficient from opening Γ_α, electrical distance from input Θ_α) and broad (respectively β, Γ_β, Θ_β) walls. One waveguide of such an array is shown in Fig. 1. For definiteness, we assume $\Theta_\alpha < \Theta_\beta$).

The structure of a scatter matrix S'' of this infinite waveguide array is the same as that of the scatter matrix S' of an infinite ray of emitters but only blocks $S''_{I II}$ and $S''_{II I}$ are not diagonal. We immediately write down the elements of the scatter matrix K of a unified waveguide element of the compensation circuit:

$$k_{11} = -\Gamma_\alpha e^{i2\Theta_\alpha} - \Gamma_\beta e^{i2\Theta_\beta} + 2\sum_{k=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k e^{i2\Theta_\alpha} + 2\sum_{k=1}^{\infty}\left(\frac{i\beta}{2}\right)^k e^{i2\Theta_\beta} +$$

$$+ 4\sum_{k=1}^{\infty}\sum_{m=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k\left(\frac{i\beta}{2}\right)^m e^{i2\Theta_\beta};$$

$$k_{22} = -\Gamma_\alpha e^{i2(\Theta-\Theta_\alpha)} - \Gamma_\beta e^{i2(\Theta-\Theta_\beta)} + 2\sum_{k=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k e^{i2(\Theta-\Theta_\alpha)} +$$

$$+ 2\sum_{k=1}^{\infty}\left(\frac{i\beta}{2}\right)^k e^{i2(\Theta-\Theta_\beta)} + 4\sum_{k=1}^{\infty}\sum_{m=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k\left(\frac{i\beta}{2}\right)^m e^{i2(\Theta-\Theta_\alpha)};$$

$$k_{12} = e^{i\Theta}\left[\sqrt{1-(\Gamma_\alpha+\Gamma_\beta)^2-\gamma^2} + 2\sum_{k=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k + \right.$$

$$\left. + 2\sum_{k=1}^{\infty}\left(\frac{i\beta}{2}\right)^k + 4\sum_{k=1}^{\infty}\sum_{m=1}^{\infty}\left(\frac{i\alpha}{2}\right)^k\left(\frac{i\beta}{2}\right)^m\right].$$

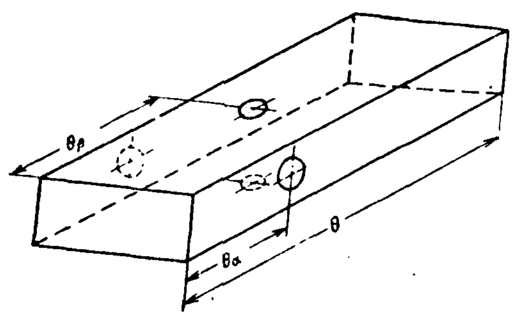

Fig. 1.

In summing the series and disregarding the small value $(\Gamma_\alpha + \Gamma_\beta)^2$, we obtain

$$k_{11} = \left(-\Gamma_\alpha + \frac{2i\alpha}{2-i\alpha}\right)e^{i2\Theta_\alpha} + \left[-\Gamma_\beta + \frac{2i\beta}{2-i\beta} - \frac{4\alpha\beta}{(2-i\alpha)(2-i\beta)}\right]e^{i2\Theta_\beta};$$

$$k_{22} = \left[-\Gamma_\alpha + \frac{2i\alpha}{2-i\alpha} - \frac{4\alpha\beta}{(2-i\alpha)(2-i\beta)}\right]e^{i2(\Theta-\Theta_\alpha)} +$$

$$+ \left(-\Gamma_\beta + \frac{2i\beta}{2-i\beta}\right)e^{i2(\Theta-\Theta_\beta)};$$

$$k_{12} = \left[1 - \gamma^2 + \frac{2i\alpha}{2-i\alpha} + \frac{2i\beta}{2-i\beta} - \frac{4\alpha\beta}{(2-i\alpha)(2-i\beta)}\right]e^{i\Theta}.$$

The value of γ^2 expresses the fraction of energy of the incident wave which departs for connection with other waveguides and can be determined from the condition of unitariness of the scatter matrix of an infinite array of waveguides which expresses the energy balance in the system

$$\gamma^2 = 16\frac{\alpha^2 + \beta^2}{(4 + \alpha^2)(4 + \beta^2)}.$$

It is easy to show that in the case of small openings

$$\sqrt{1-\gamma^2} \simeq 1 - \frac{\alpha^2 + \beta^2}{2}.$$

Thus, all elements of the scatter matrix of an element of an infinite waveguide array are expressed by the parameters of the connection opening and the length of the waveguide segments which comprise the array.

Proceeding to the transmission matrix of a waveguide in the array, we have

$$T'' = \begin{bmatrix} \frac{k_{21}k_{12} - k_{22}k_{11}}{k_{12}} & \frac{k_{22}}{k_{12}} \\ -\frac{k_{11}}{k_{12}} & \frac{1}{k_{12}} \end{bmatrix}.$$

Determination of the Parameters of the Compensation Circuit

Let us connect the unified elements of the array of emitters and array of waveguides which have been introduced with each other. Here, the compound transmission matrix of the connection is determined as

$$T = T'T''.$$

It is easy to show that the elements of the compound transmission matrix have the form

$$
\left.\begin{aligned}
T_{11} &= \frac{\left[\left(|R_{12}|^2 + |R_{11}|^2\right)\left(k_{12}^2 - k_{11}k_{22}\right) + R_{11}^* k_{11}\right]}{|R_{12}|\, k_{12}}\, e^{i\varphi_{12}} \\
T_{12} &= \frac{\left(|R_{12}|^2 + |R_{11}|^2\right) k_{22} - R_{11}^*}{|R_{12}|\, k_{12}}\, e^{i\varphi_{12}} \\
T_{21} &= \frac{-R_{11}\left(k_{12}^2 - k_{11}k_{22}\right) - k_{11}}{|R_{12}|\, k_{12}}\, e^{-i\varphi_{12}} \\
T_{22} &= \frac{-R_{11}k_{22} - 1}{|R_{12}|\, k_{12}}\, e^{-i\varphi_{12}}
\end{aligned}\right\} \qquad (3)
$$

The compound scatter matrix of the connection

$$
S = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix}.
$$

Obviously equation

$$
s_{11} = 0, \qquad (4)
$$

can serve as the condition for the compensation for the reverse interconnection, and the equation

$$
s_{22} = 0. \qquad (5)
$$

has the condition for compensation for direct interconnection.

Moving from T to S and presenting (3) in (4) and (5), we obtain the expanded conditions for the compensation of reverse and direct interconnection:

$$
R_{11}\left(k_{12}^2 - k_{11}k_{22}\right) + k_{11} = 0; \qquad (6)
$$

$$
\left(|R_{12}|^2 + |R_{11}|^2\right) k_{22} - R_{11}^* = 0, \qquad (7)
$$

where R_{11} and R_{12} are determined from (1) and (2).

We make several transformations of the elements of the matrix K, having in mind the smallness of the openings which connect the waveguides. We can assume that the transmission coefficient of the opening and the coefficient of reflection from it are connected with each other approximately in the same way as in the system of two connected waveguides [4], i.e.,

$$\alpha^2 \cong 2\Gamma_\alpha;\ \beta^2 \approx 2\Gamma_\beta.$$

In addition, in all transformations we disregard the powers of α and β above two. Obviously, we can consider the electrical length of the bridge Θ to be any one without any limitation on the generality of consideration. For example, let $\Theta = 2\pi$. Then we obtain:

$$\left.\begin{aligned} k_{11} &= \left[-\frac{\alpha^2}{2} + \frac{2i\alpha}{2-i\alpha}\right] e^{i2\theta_\alpha} + \left[-\frac{\beta^2}{2} + \frac{2i\beta}{2-i\beta} - \frac{4\alpha\beta}{(2-i\alpha)(2-i\beta)}\right] e^{i2\theta_\beta} \\ k_{22} &= \left[-\frac{\alpha^2}{2} + \frac{2i\alpha}{2-i\alpha} - \frac{4\alpha\beta}{(2-i\alpha)(2-i\beta)}\right] e^{-i2\theta_\alpha} + \left[-\frac{\beta^2}{2} + \frac{2i\beta}{2-i\beta}\right] e^{-i2\theta_\beta} \\ k_{12} &= 1 - \frac{\alpha^2+\beta^2}{2} + \frac{2i\alpha}{2-i\alpha} + \frac{2i\beta}{2-i\beta} - \frac{4\alpha\beta}{(2-i\alpha)(2-i\beta)} \end{aligned}\right\} \quad (8)$$

Solving the system of equations (6), (7), where k_{11}, k_{22}, k_{12} are determined respectively from equations (8) relative to α, β, Θ_β, we obtain:

$$\left.\begin{aligned} &\operatorname{tg} 2\Theta_\alpha = \frac{-\operatorname{Re} R_{11}}{\operatorname{Im} R_{11}};\ \operatorname{tg} 2\Theta_\beta = \frac{\operatorname{Re} R_{11}}{-\operatorname{Im} R_{11}} \\ &\alpha = 0{,}19\,(1 - |R_{11}|^2 - |R_{12}|^2);\ \beta = 0{,}31\,(1 - |R_{11}|^2 - |R_{12}|^2) \end{aligned}\right\} \quad (9)$$

Conclusion

In an infinite plane array, in contrast to a linear one, the simultaneous compensation for leakage and reradiation is possible. As is evident from formulas (9), the parameters of the compensating

openings in narrow and broad walls are different - the values of the electrical distance from the connection opening to the aperture θ_α and θ_β differ from one another by $\pi/2$, and the transmission coefficients of the openings α and β differ approximately 1.6-fold.

The formulas which have been obtained can be used to accomplish the complete decoupling of emitters in large antenna arrays.

In conclusion, I consider it my pleasant duty to express my profound graditude to O. G. Vendik for constant interest in the work and valuable remarks.

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